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MODULATION WAVEFORMS FOR CONTINUOUS WAVE RADAR

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ABSTRACT

The range of a radar is a function of average transmitter power and is independent of the modulation waveform. It does not matter whether the transmitter operates in a pulsed or CW mode. In many applications, an average power of only a few watts is adequate. In these cases, CW operation can have a number of advantages. Solid state power sources can readily be utilized and a single RF source can serve as both transmitter and local oscillator. Another advantage is waveform flexibility. In a CW radar the waveform can be designed to perform a variety of functions, e.g., detect fixed and moving targets, determine sense of radial velocity and communicate with a target over a fade-free channel.

This paper describes waveforms and design techniques that can be used to achieve the above mentioned functions with a CW radar.

SIMPLE CW RADAR

A simple CW radar utilizing a common antenna for transmit and receive is shown in Figure 1. (For simplicity, in Figure 1 and all subsequent block diagrams, no amplifiers are shown.) The RF GENERATOR and RF DEMODULATOR are coupled to the RF ANTENNA by means of an RF CIRCULATOR. Local oscillator power for the RF DEMODULATOR is obtained by leakage through the RF CIRCULATOR or directly from the RF GENERATOR. The useful output of the RF DEMODULATOR will be doppler frequency signals from moving targets within the radar beam.

A radar as shown in Figure 1 is not practical for most applications. Transmitter and mixer diode noise close to the RF carrier will fall within the doppler frequency band, resulting in poor receiver sensitivity. Since it is not possible to employ

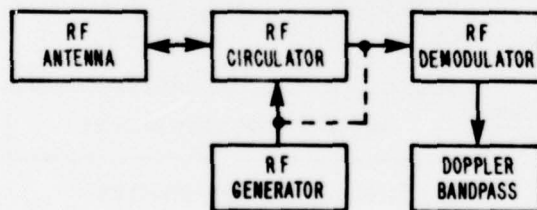


FIG. 1: SIMPLE CW RADAR

conventional sensitivity time control in the receiver, strong signals from nearby targets will mask weak returns from distant targets. Also, it is not possible to determine the range of targets. Finally, fixed target detection is impractical, since highly stable dc amplifiers would be required.

In the next several sections we will show how the above limitations can be overcome by frequency and phase modulation of the transmitter. First we will discuss solution of problem of masking by nearby targets.

SINE WAVE FM

Target output signal strength can be made almost independent of range (over certain limits) by frequency modulating the transmitter with a sine wave and processing received signals, as shown in Figure 2a. Again, the RF GENERATOR is coupled to the RF ANTENNA via the RF CIRCULATOR. Some of the transmitted signal leaks through the RF CIRCULATOR into the RF DEMODULATOR and provides a reference for the detection of returned echoes. The output of the RF DEMODULATOR is given by (Refs. 1,2).

$$E_{fm} = \sqrt{A} \left( \frac{T}{\tau} \right)^2 [E_e(t) \cos(2\pi f_0 \tau) - E_0(t) \sin(2\pi f_0 \tau)] = E_I - E_Q \quad (1)$$

where A = constant representing fixed factors in the radar range equation

$$T = \frac{1}{f_m} = \text{modulation freq period} \quad (2)$$

$\tau$  = round trip delay time of target signal

$$E_e(t) = J_0(m_f) - 2 [J_2(m_f) \cos(2\theta(t)) + \dots] \quad (3)$$

$$E_0(t) = 2 [J_1(m_f) \cos(\theta(t)) - J_3(m_f) \cos(3\theta(t)) + \dots] \quad (4)$$

$J_n$  = Bessel function of first kind, order n

$$m_f = \frac{2 \Delta f}{f_m} \sin(\pi \tau f_m) \quad (5)$$

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$\Delta f$  = peak frequency deviation of transmitter

$$\theta(t) = 2\pi f_m t - \pi f_m \tau \quad (6)$$

$f_0$  = center frequency of transmitter

$E_I$  = In-phase component of demodulated RF

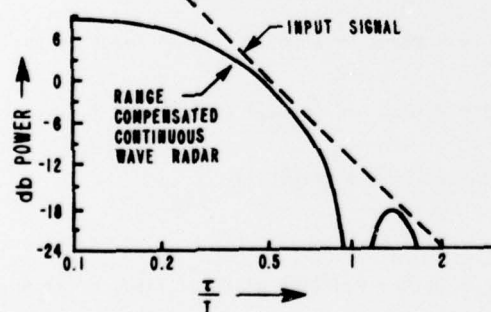
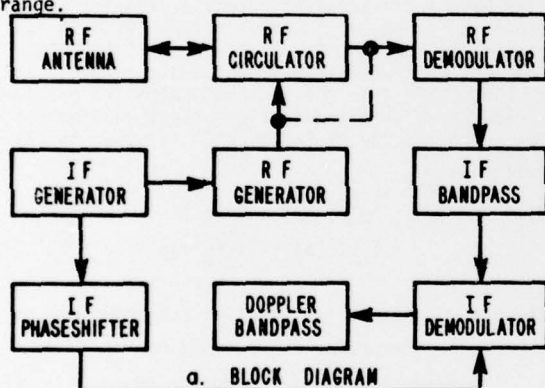
$E_Q$  = Quadrature-phase component of demodulated RF

The IF BANDPASS is tuned to  $f_m$ , thus eliminating much of the low frequency system noise.

It can be shown (Ref. 1) that the output of the IF DEMODULATOR is given by

$$E_d = \sqrt{A} \left[ \left( \frac{T}{\tau} \right)^2 \sin\left(\pi \frac{\tau}{T}\right) J_1\left(\frac{2\Delta f}{f_m} \sin\left(\pi \frac{\tau}{T}\right)\right) \right] \sin(2\pi f_0 \tau) \quad (7)$$

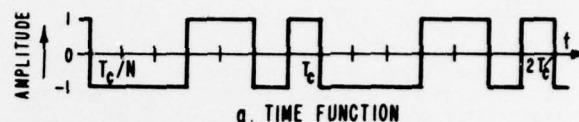
The term in brackets represents the variation of target signal output as a function of delay time. A plot of the square (proportional to power) of this function vs  $\tau/T$  is shown in Figure 2b for the case where  $2\Delta f/f_m$  is equal to 1.8. The minimum processing loss (about 2 dB) occurs when  $\tau/T$  is 0.5. A design criterion is then that the maximum target delay time be equal to  $\frac{1}{2}T$ . The signal output varies by about 10 dB as the target moves from maximum to zero range. Also shown in Figure 2b is the input signal strength, which decreases at the rate of 12 dB per octave of range.



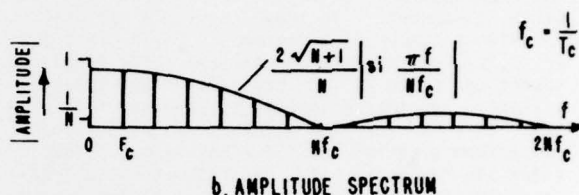
b. POWER OUTPUT VERSUS RANGE DELAY TIME

FIG. 2: CW RADAR WITH RANGE COMPENSATION

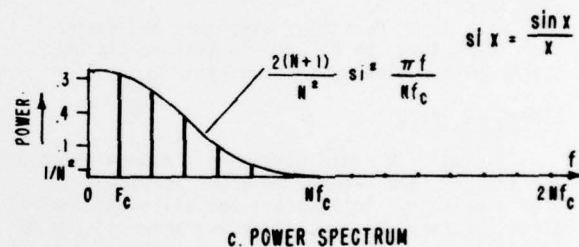
Two factors contribute to the signal compression. First, for small arguments the first order Bessel function is very nearly proportional to  $\tau$ , thereby giving a 6 dB reduction in signal each time the range is halved. Another 6 dB reduction is obtained from the  $\sin(\pi \tau/T)$  term. This term comes about through the change in phase of the IF signal with range.



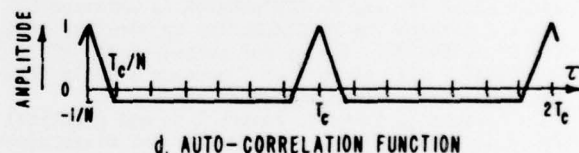
a. TIME FUNCTION



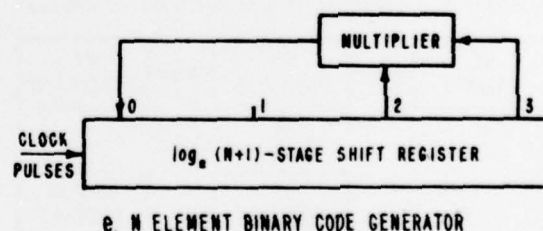
b. AMPLITUDE SPECTRUM



c. POWER SPECTRUM



d. AUTO-CORRELATION FUNCTION



e. N ELEMENT BINARY CODE GENERATOR

FIG. 3: MAXIMUM LENGTH PSEUDO RANDOM CODE

## RANGE RESOLUTION

In order to obtain range resolution in a CW radar, it is necessary to modulate the transmission. A suitable modulation waveform is a maximum length binary pseudo-random code (Refs. 3,4). A seven-element code is shown in Figure 3. Since the pseudo-random code is quantized in both amplitude and time, it can be conveniently generated by a shift register with feedback. The feedback is accomplished through binary multiplication of the outputs from two of the shift register stages, using the product as the input to the first stage. The operation of a multiplier is such that its output is positive if the inputs are of like polarity, and negative if the inputs are of opposite polarity. The code length  $N$  is  $2^n - 1$  elements, where  $n$  is the number of stages in the shift register.

A three-stage shift register with feedback from the second and third stage is shown in Figure 3e. With every clock pulse the information in each stage of the SHIFT REGISTER is transferred to the following stage. If an output is taken from one stage, a code will be generated which repeats every seven clock pulses.

From Figure 3e, it is evident that the output from the second stage of the SHIFT REGISTER is delayed by one clock period (time between successive clock pulses) from the output of the first stage. Hence, any desired amount of delay can be obtained by cascading additional stages. A delay can also be obtained by using a second shift register and delaying its clock pulses with respect to those of the first.

The code has three interesting features. If we represent the code waveform by  $u(t)$ , then when  $\tau_1$ ,  $\tau_2$ , and  $\tau_3$  are multiples of a clock period,

$$u(t) \cdot u(t) = 1 \quad (8)$$

$$u(t - \tau_1) u(t - \tau_2) = u(t - \tau_3) \quad (9)$$

$$\overline{u(t)} = \frac{1}{N} \quad (10)$$

Equation 8 says that a code multiplied by itself yields a constant value of unity, while Equation 9 says that the product of two shifted codes is again a code, but with a new shift. Equation 10 says that the mean value of the code is  $1/N$ . These three features give rise to an autocorrelation function, as indicated in Figure 3d, with a sidelobe level of  $1/N$ . For example, if  $N$  is 1023 the theoretical peak to sidelobe ratio is over 60 dB. In practice, it is difficult to meet the component tolerances required to obtain more than 50 dB.

In some applications a side lobe level of 50 dB is not adequate. A technique for increasing the rejection of off-range targets is discussed in the following section. This technique is applicable when the highest doppler frequency is less than one tenth of the code repetition frequency.

## COMBINATION OF PSEUDO-RANDOM CODE MODULATION AND SINE WAVE FM

The sine wave FM can be combined with the pseudo-random code modulation if certain precautions are taken. A block diagram of the combined system is shown in Figure 4a. The RF GENERATOR is

sine wave frequency modulated. The output of the RF GENERATOR is applied to the RF MODULATOR, where it is balanced modulated by the CODE GENERATOR. The RF MODULATOR changes the phase of the transmitted signal by  $180^\circ$  each time the code changes state. The RF MODULATOR is coupled to the RF ANTENNA via the RF CIRCULATOR. Again, transmitter leakage through the RF CIRCULATOR into the RF DEMODULATOR is the product of transmitted code times the returned code modulated by an FM waveform. The modulation imparted by the mixing of the transmitted and received signals in the RF DEMODULATOR is accounted for by multiplying the delayed code by the transmitted code in the CODE MULTIPLIER of Figure 4a. The output of the CODE MULTIPLIER is applied to the CODE DEMODULATOR. If the CODE DELAY is  $\tau'$ , the output of the CODE DEMODULATOR is then

$$E_s(t) = (E_I - E_Q) u(t - \tau) u(t - \tau') \quad (11)$$

Since the IF BANDPASS will pass only frequencies close to  $f_m$ , it is necessary to examine the spectrum of  $E_s(t)$ . The spectrum of  $E_s(t)$  is the convolution of the code spectrum with harmonics of  $f_m$ . In order to maintain the  $1/N$  peak-to-sidelobe

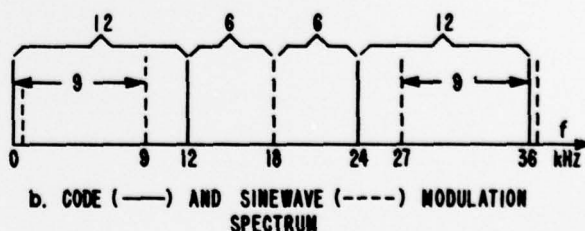
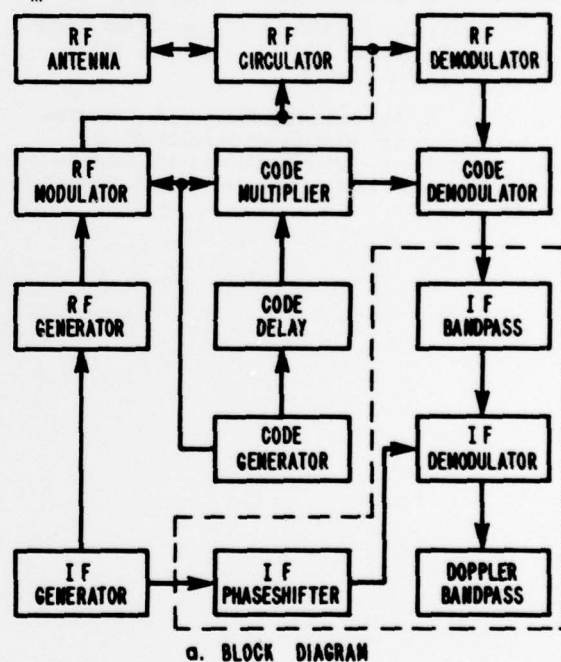


FIG. 4 : CW RADAR WITH RANGE RESOLUTION



as the target doppler frequency is below 100 cps, the speech intelligibility will not be appreciably affected.

If "a" and "b" are not equal, then the signal amplitude will change with target range, but will never be less than the smaller of "a" or "b".

#### CONCLUSIONS

The techniques we have discussed illustrate the waveform design flexibility available in a CW radar. The application and extrapolation of these techniques can improve the performance of many types of radars; e.g., surveillance, collision avoidance, marine, intrusion detection and police.

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ratio of the pseudo-random code, it is necessary that the  $f_m$  component in the spectrum of  $E_s(t)$  arise only from the product of the  $J_1$  term of  $E_0$  and the dc component of the code spectrum. As can be seen from Figure 3b, if harmonics of the code repetition frequency give rise to an  $f_m$  component in  $E_s(t)$ , the sidelobe level of the autocorrelation function will be down only by approximately  $1/\sqrt{N}$  instead of  $1/N$ .

The above mentioned requirement on the spectrum of  $E_s(t)$  can be satisfied by taking advantage of the rapid decrease in amplitude of the higher order Bessel functions. For all practical purposes the spectrum due to the frequency modulation (the spectrum of  $E_1 - E_0$ ) is finite. If we keep  $m_f$  below 1.5, then we need consider only frequencies up to  $3 f_m$ .

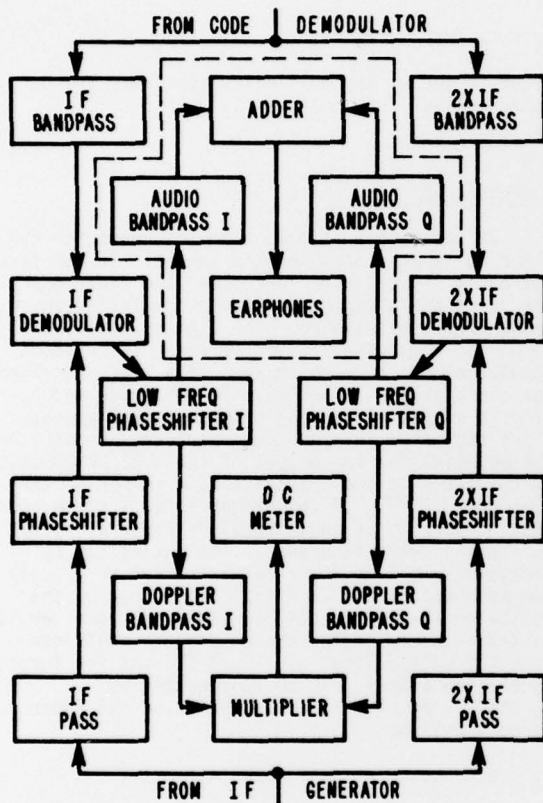


FIG. 5 : SINGLE SIDE BAND PROCESSOR

The procedure for the selection of  $f_m$  and the code repetition frequency,  $f_c$ , is best illustrated by an example. Consider the case of a radar which is required to detect targets out to a range of 10 km. A suitable  $f_c$  is 12 kHz. If  $f_m$  is 9 kHz, then an undesired 9 kHz signal can arise from a beat between the 3rd harmonic of  $f_c$  (36 kHz) and the 3rd harmonic of  $f_m$  (27 kHz). The relationship of the various frequencies is shown in Figure 4b.

The voltage ratio of the desired to undesired signals is given by the following expression:

$$R_{13} = J_1(m_f) / (\sqrt{N} J_3(m_f)) \quad (12)$$

If  $m_f$  is limited to 1.5 at maximum range, then  $R_{13}$  will be at least 35 dB even when the code length is as short as 63. As the target range decreases  $m_f$  will decrease and the  $J_3$  component will drop off rapidly. The drop off will be more rapid than the increase in returned power, due to decreasing range. Thus the rejection ratio of off-range targets will be at least 35 dB at all ranges. A 9 kHz signal can also arise from the 5th harmonic (45 kHz) of  $f_m$ , however, its amplitude will be negligible.

The choice of 9 kHz and 12 kHz limits the available doppler band to about 1000 cps. The limitation on the doppler band is due to the fact that close-in targets will produce strong code lines in the output of the CODE DEMODULATOR. If target has a doppler frequency of 1000 cps, the code will have a component of 11 kHz. For a 63 element code, this component will be down by only 18 dB from the entire signal. When the 11 kHz component mixes with the 9 kHz reference in the IF DEMODULATOR, the beat frequency will be 2 kHz. To maintain an 80 dB peak to sidelobe ratio, the 2 kHz component must be attenuated 62 dB by the DOPPLER BANDPASS. If the cut-off frequency of DOPPLER BANDPASS is 1000 Hz, the required attenuation can be obtained with a small filter. If desired, some of the attenuation of the 11 kHz signal can be effected in the IF BANDPASS.

In order to avoid any spurious signals, it is essential that all beat frequencies between oscillators in the system be a multiple of 3 kHz. For example, the power supply and the oscilloscope sweeps should operate on harmonics of 3 kHz. To lock  $f_m$  and  $f_c$ , they should be generated from the same oscillator.

#### FIXED TARGET DETECTION

In order for fixed targets to be detected by the radar of Figure 4 they must be modulated by a frequency which falls within the doppler passband. Fixed targets can be modulated by frequency modulating the RF GENERATOR with a second sine wave. If the frequency of this sine wave is  $f_f$  and the peak transmitter frequency deviation is  $\Delta f$ , it can be shown that the output of the CODE DEMODULATOR,  $E_s(t)$ , will be modified such that

$$\frac{\cos[2\pi f_0 \tau]}{\sin[2\pi f_0 \tau]}$$

is replaced by

$$\frac{\cos[2\pi f_0 \tau + \frac{2\Delta f}{f_m} \sin(\pi f_f \tau) \sin(2\pi f_f \tau - \pi f_f \tau)]}{\sin[2\pi f_0 \tau]}$$

(13)

### SENSE OF RADIAL VELOCITY

Sense of radial velocity can be determined by processing quadrature components of the RF signal. As shown by the equation for the output of the RF DEMODULATOR,  $E_s(t)$ , quadrature components of the RF signal,  $\cos(2\pi f_0 \tau)$  and  $\sin(2\pi f_0 \tau)$ , are available. The sine component appears at the output of the IF DEMODULATOR (Figure 4a). The cosine component can be obtained by processing the second harmonic of the FM waveform, i.e., the  $J_2$  component. The second harmonic is processed in essentially the same manner as the first harmonic (see Figure 5). The only difference besides using a bandpass centered at  $2f_m$  is the setting of the IF Phase Shifter. For small arguments, the second order Bessel function decreases by 12 dB each time the argument is halved. Thus, the amplitude of the  $J_2$  component of the returned signal will be nearly constant with range. The phase of the  $J_2$  component varies by  $180^\circ$  as the target delay changes from zero to  $\frac{1}{2}T$ . It is then necessary to couple the IF PHASE SHIFTER to the range delay control in order to maximize the signal.

Let the output of the IF channel be represented by

and the output of the 2 IF channel by

If the target moves,  $\tau$  will vary as a function of time. The outputs will then be

$$E_2 = K_2 \cos(\phi + 2 \pi f_d t) \quad (17)$$

$f_d$  will be positive if the target is receding and negative if the target is approaching.

After  $E_1$  and  $E_2$  are multiplied together, the DC component is

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It can be seen that the amplitude of  $E_v$  is independent of  $r$ . Changes in target range show up as a phase modulation of the voice signal. As long

on of the voice signal. As long

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